Unique compensation technique tames high-bandwidth voltage-feedback op amps

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You may think that there's nothing new in op-amp compensation techniques. Think again. A unique and previously overlooked method allows a decompensated voltage-feedback op amp to achieve low-gain operation with high dc accuracy, high slew rate, and low harmonic distortion.

Designers seeking high slew rates and low noise for dc-coupled pulse amplifiers often must turn to extremely high-gain-bandwidth, nonunity-gain-stable, voltage-feedback op amps. The lower internal compensation capacitance, which gives these op amps the nickname "decompensated," increases slew rate, and the higher input-stage transconductance, g_m, which produces the ultrahigh gain bandwidth, decreases input-voltage noise.

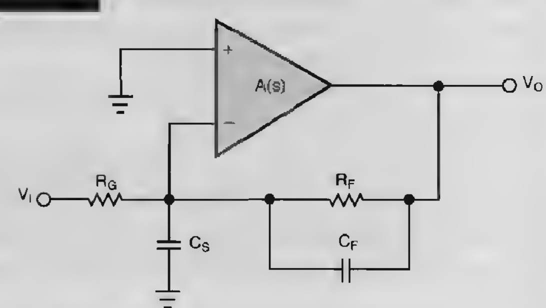
Unfortunately, many designers have been burned trying to apply these touchy decompensated devices to low gains. Much of the popularity for the current-feedback topology comes from its superior slew rate and stability at low gains compared with high-gain-bandwidth voltage-feedback designs. However, the high-frequency performance of a current-feedback op amp also comes with poor dc accuracy and higher output noise.

Op-amp designers suggest various forms of external compensation to take advantage of the dc accuracy, low noise, and high slew rate of a decompensated voltage-feedback op amp at low signal gains. Unfortunately, previously suggested compensation schemes have many shortcomings. For example, some op amps provide access to the internal compensation node, but adding this dominant-pole compensation directly reduces the slew rate. Common lead-lag compensation techniques produce pole-zero pairs in the closed-loop response, yielding deplorable pulse response and settling characteristics.

A new external compensation method provides complete control over a simple, second-order lowpass response at low signal gains. This technique allows you to achieve a well-controlled frequency response at any inverting gain for any internally decompensated op amp. The full slew rate of the decompensated op amp is available at the output, along with an output-noise voltage density that increases with frequency. This increased output noise stems from the necessary peaking in the noise gain to achieve a flat, closed-loop frequency response. Passive postfiltering can significantly reduce the effect of this noise.

Using this external technique with a high-quality, decompensated voltage-feedback op amp provides significantly better absolute dc accuracy than high-speed current-feedback alternatives. Comparable noise and slew rate

FIGURE 1



NOTES:
$A(s)=\frac{A_{OL}\omega_A}{s+\omega_A}$.
A_{OL}=OPEN-LOOP GAIN.
ω_A=RADIAN DOMINANT POLE.
$\frac{A_{OL}\omega_A}{2\pi}$=GAIN-BANDWIDTH PRODUCT (GBP).

A simple but previously unexplored compensation circuit consists of C_F and C_S. The technique allows you to use a decompensated voltage-feedback op amp at low gains with the high-frequency benefits of a current-feedback op amp but the dc accuracy of the voltage-feedback part.

and considerably lower harmonic distortion than equivalent current-feedback options are also possible. With some extra effort, you can also use this compensation to emulate the gain-bandwidth independence of a current-feedback op amp. Gain-bandwidth independence using a voltage-feedback op amp can be useful in inverting-summing applications for which you might need to adjust the summing weights during the design process or as part of the application.

Once you understand the topology and derive the basic transfer function, you can predict the amplifier's performance based on the desired signal gain and the amplifier's characteristics. Three design examples show how the compensation technique works to maximize the achievable flat bandwidth, implement a filter, or produce a gain-bandwidth-independent design (and why you would want that).

Analyze the compensation circuit

The compensation technique simply consists of adding two compensation elements, C_S and C_F, to the standard inverting op-amp configuration (**Figure 1**). Previous discussions of this circuit focused on using C_F to compensate for a parasitic C_S. The following analysis shows you how to set both C_S and C_F to get a well-controlled, closed-loop, second-order lowpass frequency response at any signal gain for even the most unstable op amp.

You can easily analyze this circuit using a single-pole,

SECOND-ORDER LOWPASS-RESPONSE CHARACTERISTICS

A good understanding of the second-order lowpass transfer function is important to understanding this compensation technique. **Equation A** shows the general form of the Laplace transfer function of a second-order lowpass response:

$$\frac{V_O}{V_I}=\frac{A\omega_0^2}{s^2+s\frac{\omega_0}{Q}+\omega_0^2}. \quad \textbf{(A)}$$

The characteristic frequency is the radial distance in the s-plane from the origin to the poles when they are complex-conjugate pairs. The units in **Equation A** normally give this frequency in radians; you can convert values to hertz by dividing by 2π. The Q indicates how complex the poles are. The angle that the vector makes with the negative-real axis in the s-plane from the origin to the complex poles is given by $\cos^{-1}(1/2Q)$. Some key values for Q are

- when Q<0.5, the poles are both real;
- when Q=0.5, two repeated real poles occur at $-\omega_0$;
- when Q=0.577, the frequency response is a second-order Bessel with the best phase linearity; and
- when Q=0.707, the frequency response is a second-order Butterworth with maximum flatness.

At Q=0.707, the poles are at ±45° to the negative-real axis in the s-plane. The pulse response for a Butterworth lowpass response shows about 4.3% overshoot. For Q>0.707, the frequency response begins to peak, extending the –3-dB bandwidth but also introducing additional overshoot in the step response.

Another interesting point for Q is the geometric mean of the Bessel and Butterworth Qs. This point gives a Q=0.639, which is attractive because it gives the highest bandwidth (for a given ω_0) with less than 2% overshoot in the pulse response.

You can describe the –3-dB bandwidth as a function of ω_0 and Q as follows:

$$F_{-3\,dB}=\frac{\omega_0}{2\pi}\sqrt{\left(1-\frac{1}{2Q^2}\right)+\sqrt{\left(1-\frac{1}{2Q^2}\right)^2+1}}. \quad \textbf{(B)}$$

Evaluating the radical portion of **Equation B** at several of the possible values of Q suggested above gives the ratio of the –3-dB bandwidth to the characteristic frequency.

For Q=0.577, $F_{-3\,dB}=0.79F_0$. For Q=0.639, $F_{-3\,dB}=0.90F_0$. For Q=0.707, $F_{-3\,dB}=F_0$.

Some other useful second-order lowpass-response **equations** include

$$F_{PEAK}=\frac{\omega_0}{2\pi}\sqrt{1-\frac{1}{2Q^2}},$$

which is the peak frequency when Q>0.707;

$$\text{GAIN PEAKING}=20\log\left[\frac{Q}{\sqrt{1-\frac{1}{4Q^2}}}\right]\text{dB},$$

which is the amount of peaking when Q>0.707;

$$\text{PERCENT OVERSHOOT}=100\%\left[e^{-\pi\cot\left(\cos^{-1}\frac{1}{2Q}\right)}\right],$$

which is the percent overshoot in the pulse response when Q>0.50; and

$$\text{NPB}=\frac{\omega_0}{2\pi}\left[\frac{\pi}{2}Q\right](\text{Hz})=F_0\frac{\pi}{2}Q(\text{Hz}),$$

which is the noise-power bandwidth when Q>0.5.

open-loop model for the op amp. Without C_S and C_F, a single-pole op-amp model would be inadequate because the higher order poles of a decompensated op amp wholly determine the closed-loop response at low gains. However, you'll see that the design methodology justifies this single-pole simplification with the compensation elements in place.

Besides being the only way this compensation will work, the inverting configuration offers several other benefits. With no common-mode voltage at the input, the inverting configuration for most op amps achieves higher slew rates, higher full-power bandwidth, and lower distortion. The trade-offs to getting these inverting-mode benefits are an input impedance set by R_G and a slightly higher dc noise gain for the noninverting input-voltage noise of the op amp.

You can write the Laplace transfer function for the circuit of **Figure 1** in Bode-analysis form as follows:

$$\frac{V_O}{V_I}=\frac{-Z_F/R_G}{1+\left(\frac{1+Z_F/Z_G}{A(s)}\right)}, \quad (1)$$

where

$$Z_F=R_F \,||\, \frac{1}{sC_F}=\frac{1/C_F}{s+1/R_FC_F}, \quad (2)$$

$$Z_G=R_G \,||\, \frac{1}{sC_S}=\frac{1/C_S}{s+1/R_GC_S}, \quad (3)$$

TABLE 1—OP-AMP SPECIFICATIONS

Typical specifications	OPA627	OPA637
GBP (MHz)	16	80
Minimum stable gain	1	5
Slew rate (V/μsec)	55	135
Input-voltage noise (nV/$\sqrt{Hz}$)	4.5	4.5
Input capacitance ($C_{differential}$+ $C_{common\ mode}$) (pF)	15	15
Input-offset voltage (mV)	0.1 (low grade)	0.28 (low grade)

and the single-pole op amp's open-loop gain is

$$A(s)=\frac{A_{OL}\omega_A}{s+\omega_A}. \quad (4)$$

The significant components of this transfer function are

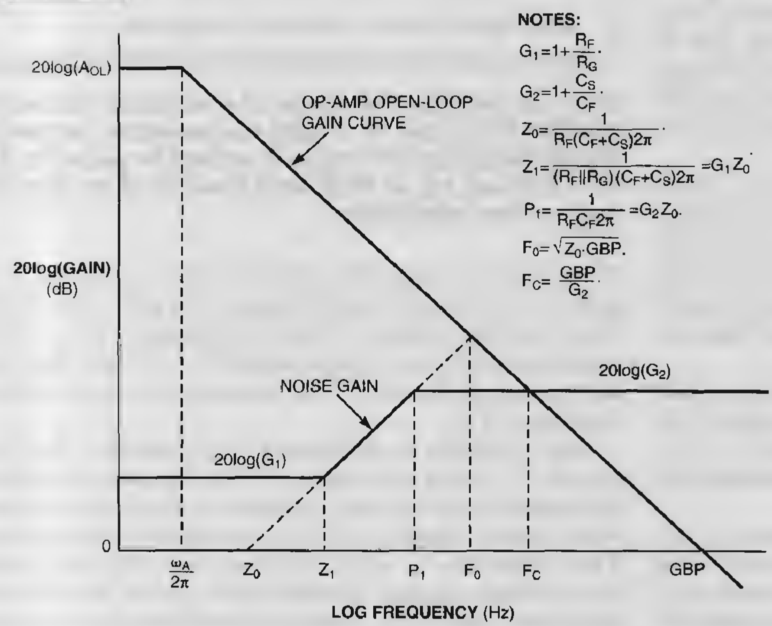
1. $-Z_F/R_G$, which would be the signal gain if the op amp were ideal (had infinite open-loop gain and bandwidth),
2. $1+Z_F/Z_G$, which is the noise-gain portion of the loop gain (and also equal to the gain from the noninverting input to the output), and
3. A(s), which is the open-loop gain over frequency for the op amp.

At dc, the denominator of **Equation 1** is approximately 1, whereas the numerator is equal to $-R_F/R_G$, which is the desired low-frequency signal gain. For stability analysis, it is common to look at the corresponding Bode plot (**Figure 2**).

The magnitude portion of the Bode plot compares the magnitude of the noise gain with the magnitude of the open-loop gain, which are the top and bottom of the fraction in the denominator of **Equation 1**, respectively. At the frequency at which these two curves cross, which is loop-gain crossover, the loop gain drops to 1 and, in a simple op-amp application, the closed-loop bandwidth rolls off. Because a pole also exists in the numerator of **Equation 1**, this simple analysis is not sufficient to determine the closed-loop response.

Normally, you would also need to consider the phase of the loop-gain terms. However, **Equation 1** ultimately reduces to a simple, second-order

FIGURE 2



For the compensation-network design, Bode analysis points out the unity-gain intersection of the sloping portion of the noise-gain curve (Z_0), the pole set by the feedback-compensation network (P_1), the low-frequency noise gain (G_1), and the noise gain at loop-gain crossover (G_2).

OP-AMP COMPENSATION

lowpass transfer function, and you proceed with the design by controlling the ω_0 and Q of that transfer function. The magnitude portion of the Bode analysis provides insight into what is happening in the design, but you don't use the magnitude information to set C_S and C_F. You can disregard the phase plot for now with the assumption that loop-gain crossover will occur at a noise gain high enough for you to safely ignore the higher order poles of A(s).

Substituting the two impedances, Z_F and Z_G, and the op amp's open-loop-gain expression A(s) into **Equation 1** yields

$$\frac{V_O}{V_I}=\frac{-\frac{1}{R_GC_F}\,\frac{1}{\left(s+\frac{1}{R_FC_F}\right)}}{1+\dfrac{\left[1+\dfrac{\frac{1}{C_F}}{\left(s+\frac{1}{R_FC_F}\right)}\left(\dfrac{s+\frac{1}{R_GC_S}}{\frac{1}{C_S}}\right)\right]}{\left(\dfrac{A_{OL}\omega_A}{s+\omega_A}\right)}}. \qquad (5)$$

Rearranging this **equation** to produce a pole-zero expression for the noise-gain terms in the denominator yields

$$\frac{V_O}{V_I}=\frac{-\frac{1}{R_GC_F}\,\frac{1}{\left(s+\frac{1}{R_FC_F}\right)}}{1+\dfrac{\left[\dfrac{\left(1+\frac{C_S}{C_F}\right)\left(s+\frac{1}{(R_F\|R_G)(C_S+C_F)}\right)}{\left(s+\frac{1}{R_FC_F}\right)}\right]}{\left(\dfrac{A_{OL}\omega_A}{s+\omega_A}\right)}}. \qquad (6)$$

The terms in the denominator make up the loop-gain portion of this transfer function. The op amp's open-loop gain has a high dc value of A_{OL} and a dominant pole at ω_A. The noise gain has a dc gain of $1+R_F/R_G$, a low-frequency zero, and a high-frequency pole to flatten the noise gain to $1+C_S/C_F$ at higher frequencies. The complete Bode plot (**Figure 2**) shows the gain-magnitude portion for this loop gain along with a number of key frequencies that are critical to the design.

The key frequencies (in hertz) are GBP, Z_0, and P_1. GBP is simply the gain-bandwidth product of the selected op amp ($GBP=A_{OL}\omega_A/2\pi$ Hz). Z_0, which equals $1/(2\pi R_F(C_S+C_F))$, is the unity-gain (0-dB) intersection of the sloping portion of the noise-gain curve. The actual zero in the noise gain occurs at $G_1Z_0=Z_1$. G_1 and G_2 are the low-frequency and high-frequency noise gains, respectively.

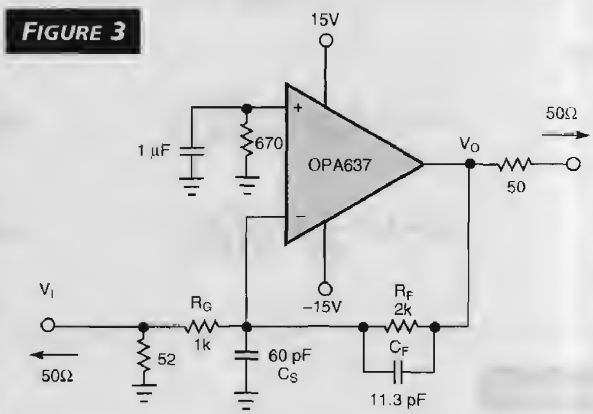
P_1, the feedback-network pole, is equal to $1/(2\pi R_FC_F)$. This pole and Z_0 are the two things you can adjust to control the closed-loop frequency response. P_1 is also equal to Z_0G_2, which is simply Z_0 times the high-frequency noise gain set by the capacitor ratios.

Another point of interest from **Figure 2** is where the projection of the sloping portion of the noise-gain curve intersects the open-loop-gain curve at the geometric mean of Z_0 and GBP. This point turns out to be the characteristic frequency, F_0, of the closed-loop second-order response (see box, "Second-order lowpass-response characteristics").

TABLE 2—OPA637 GAIN OF −2 EXAMPLE

Design features	Values	Notes
Target Q	0.64	See **box**
Target G_2	7.5	1.5 times the minimum stable gain
Computed Z_0	0.93 MHz	From **Equation 13**
Resulting Z_1	2.77 MHz	$G_1 \cdot Z_0$
Resulting P_1	6.93 MHz	$G_2 \cdot Z_0$
Resulting F_0	8.6 MHz	$\sqrt{Z_0 \cdot GBP}$
Resulting $F_{-3\,dB}$	7.74 MHz	if Q=0.64
Selected R_F	2 kΩ	
Required R_G	1 kΩ	
Required C_F	11.5 pF	From **Equation 15**
Required C_S	75 pF	From **Equation 16**

FIGURE 3



NOTE: FIGURE DOES NOT SHOW SUPPLY DECOUPLING.

Once you complete the design of a maximum-bandwidth amplifier with a gain of −2 using compensation, you have to slightly modify capacitor values to account for parasitics. A C_F of 11.3 pF and a C_S of 60 pF are 0.2 and 15 pF less than the design values, respectively.

When you set P_1 to less than this geometric mean, the noise gain crosses the open-loop response at a gain equal to G_2. The noise gain crosses the open-loop response at F_C, which would equal the closed-loop bandwidth for a unity-gain-stable op amp of the same GBP operating at a noninverting noise gain of G_2.

One of the key assumptions in this analysis is that you control G_2 so that it's greater than the specified minimum stable gain for the op amp. Crossover at this high noise gain is the reason you can use a nonunity-gain-stable op amp at a low signal gain of $-R_F/R_G$. There is, however, little consistency among op-amp manufacturers on the definition of minimum stable gain. Some manufacturers use a typical phase-margin target, others target a maximum peaking, and still others actually specify a gain that causes oscillation in

the closed-loop response. Generally, most data sheets show a recommended minimum gain that does not cause oscillation. The goal in this design is for the noise gain to cross over the open-loop response at a noise gain, G_2, high enough for you to safely ignore the higher order poles of A(s). If the minimum stable gain on the data sheet is really a minimum operating suggestion, it should be safe to target crossover at 1.5 times that gain. This guard band is, however, an estimate and varies from part to part and from manufacturer to manufacturer. Using the macromodels that most manufacturers provide allows you to fine-tune this target.

You can extensively use the frequencies and gains in the Bode plot to gain insight into the algebraic solution for the closed-loop, second-order transfer function. Because the design seeks values for the compensation elements (C_F and C_S), the following methodology uses radian frequency units. Converting those units to the hertz shown in **Figure 2** simply requires a division by 2π.

Expanding the transfer function of **Equation 6** into normal monic form (writing a polynomial from highest order to lowest order with a coefficient of 1 for the highest order term) yields

$$\frac{V_O}{V_I} = \frac{-\dfrac{A_{OL}\omega_A}{(C_F+C_S)R_G}}{s^2 + s\dfrac{\omega_0}{Q} + \omega_0^2}, \quad (7)$$

where

$$\omega_0 = \sqrt{\frac{\omega_A}{(C_F+C_S)R_F}\left(A_{OL}+1+\frac{R_F}{R_G}\right)}, \quad (8)$$

TABLE 3—LOWPASS-FILTER DESIGN

Design targets	Resulting design	Component values (R_F=2 kΩ)
ω_0=5 MHz	Z_0=312 kHz	R_G=1 kΩ
Q=2	G_2=51.2	C_F=5 pF
G_1=3	P_1=16 MHz	C_S=249 pF

and

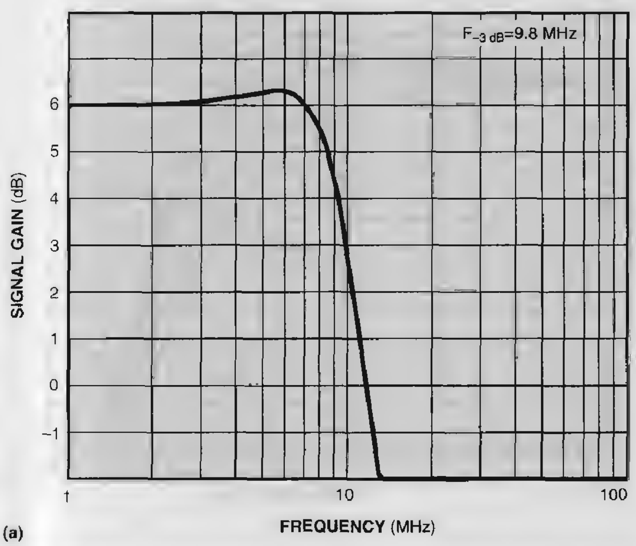
$$Q = \frac{\sqrt{\dfrac{\omega_A}{(C_F+C_S)R_F}\left(A_{OL}+1+\dfrac{R_F}{R_G}\right)}}{\dfrac{\omega_A}{(C_F+C_S)}\left(C_F(A_{OL}+1)+C_S\right)+\dfrac{1}{(R_F \parallel R_G)(C_F+C_S)}}. \quad (9)$$

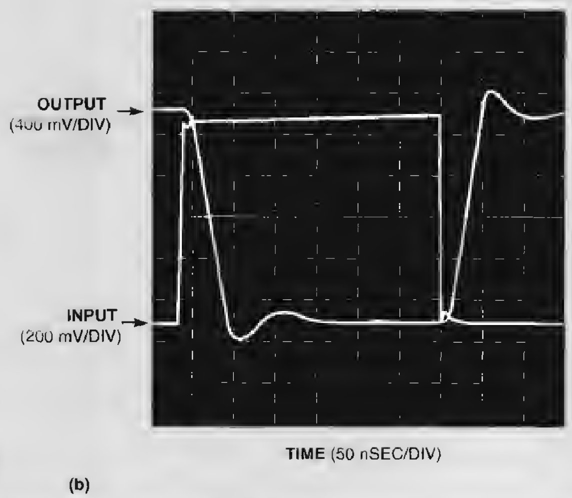
Although seeing that this full transfer function ends up as a second-order lowpass response is encouraging, the individual terms still look a little intractable. With a bit of manipulation and judicious simplifications, you can develop simple expressions for ω_0 and Q that show a clear path to a design methodology.

Specifically, you can simplify the terms inside the radical for ω_0 by recognizing that A_{OL} is much greater than $1+R_F/R_G$. Dropping the $1+R_F/R_G$ of that term, recognizing that $A_{OL}\omega_A$=GBP and that $1/(C_F+C_S)R_F=Z_0$ (in **Figure 2**), and simplifying the expression for Q in the denominator yields the following **equations**, where $G_2=1+C_S/C_F$ and $G_1=1+R_F/R_G$:

$$\omega_0 = \sqrt{Z_0 GBP} = 2\pi F_0, \quad (10)$$

FIGURE 4





The small-signal frequency response (a) of the maximum-bandwidth design shows a remarkably flat response for an op amp that is not intended for use at low gains. The response also includes slightly more peaking than expected, which introduces some overshoot and ringing into the pulse response of a ±1V output swing (b).

and

$$Q = \frac{\sqrt{Z_0 GBP}}{\frac{GBP}{G_2} + G_1 Z_0}. \quad \textbf{(11)}$$

Referring back to the Bode plot of **Figure 2**, these simple **equations** indicate that the closed-loop, second-order response has a characteristic frequency, ω_0, that is the geometric mean of Z_0 and the amplifier's GBP. Also, the ratio of that characteristic frequency to the sum of the high-frequency, loop-gain crossover frequency (F_C) and the zero frequency in the noise gain (Z_1) sets the value of Q. If you've already selected the amplifier and the required signal gain (G_1=|SIGNAL GAIN|+1), you need only set Z_0 and P_1 (or, equivalently, G_2), to implement the compensation.

Design for maximum bandwidth

Virtually all the elements that determine the Q of the closed-loop response in **Equation 11** are known. The system designer determines the amplifier's GBP and the desired low-frequency noise gain. Once you select a target Q, you need only set Z_0 and G_2. The key simplification to this analysis is to judiciously target a G_2 that is greater than the specified minimum stable gain for the selected op amp so that you can continue to neglect the added phase shift that the high-frequency, open-loop poles introduce. To get as much bandwidth as possible, set the target G_2 very close to the minimum stable gain. As previously suggested, the following design examples use a factor of 1.5 times the minimum stable gain. With G_2 somewhat arbitrarily set, you can then use **Equation** 11 to solve for Z_0. The following **equation** shows the solution as a quadratic equation that you must solve to set Z_0:

$$Z_0^2 - Z_0 GBP\left[\frac{1}{(Q \bullet G_1)^2} - \frac{2}{G_1 G_2}\right] + \left(\frac{GBP}{G_1 G_2}\right)^2 = 0. \quad \textbf{(12)}$$

An exact solution for Z_0 is

$$Z_0 = \frac{GBP}{2}\left[\frac{1}{(Q \bullet G_1)^2} - \frac{2}{G_1 G_2}\right]\left[1 - \sqrt{1 - \frac{4}{\left(\frac{1}{Q^2} \bullet \frac{G_2}{G_1} - 2\right)^2}}\right]. \quad \textbf{(13)}$$

However, when $(G_2/G_1) > 6Q^2$, a good approximation is

$$Z_0 \approx \frac{GBP \bullet Q^2}{G_2\left[G_2 - 2Q^2 G_1\right]}. \quad \textbf{(14)}$$

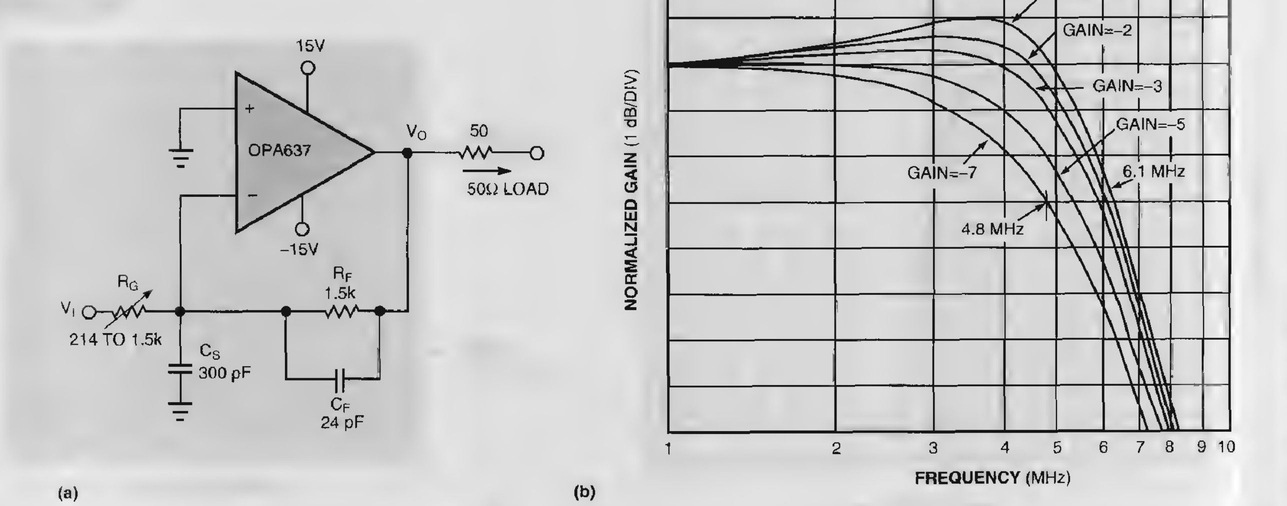
After selecting G_1 and G_2 and determining Z_0, you can implicitly determine P_1: $P_1 = G_2 Z_0$. Then, you can combine the **equations** for Z_0 and G_2 in **Figure 2** to solve for C_F and C_S:

$$C_F = \frac{1}{R_F Z_0 G_2}, \quad \textbf{(15)}$$

and

$$C_S = (G_2 - 1)C_F. \quad \textbf{(16)}$$

FIGURE 5



Applying the compensation technique to achieve gain-bandwidth independence allows you to use the voltage-feedback op amp for an adjustable-gain, constant-bandwidth circuit (a). Adjusting the gain from −1 to −7 decreases the bandwidth by only 21%, from 6.1 to 4.8 MHz (b).

Note that with a target Q of 0.707, you can substitute **Equation 13** into **Equation 10** to show the maximum achievable F_0, which is approximately equal to $F_{-3\,dB}$ given the following factors: a given op amp's GBP, the G_1 that corresponds to the desired signal gain, and the high-frequency gain, G_2, necessary for stability. The resulting **equation** shows the maximum achievable flat bandwidth using this compensation technique:

$$F_{0(MAX)} \approx \frac{GBP}{\sqrt{2}}\sqrt{\frac{1}{G_2(G_2-G_1)}}\Bigg|_{Q=0.707} \approx F_{-3\,dB}. \qquad (17)$$

Maximize the bandwidth of a real design

One of the greatest attractions of this compensation technique is that it allows you to successfully apply a nonunity-gain-stable, voltage-feedback op amp at low signal gains and simultaneously retain the full slew rate and dc accuracy of the part. **Table 1** summarizes the key specifications for a pair of good voltage-feedback op amps. The OPA627 from Burr-Brown Corp is unity-gain-stable; the OPA637 is the company's decompensated version and has a recommended minimum gain of 5. In this case, the input-voltage noise of the decompensated OPA637 is no lower than that of the OPA627, but the slew rate (and high-frequency open-loop gain) of the OPA637 is markedly higher than that of the OPA627. You can compare the OPA627 performance to the performance of the OPA637 in a complete design that uses the compensation technique.

Table 2 summarizes a gain of –2 (G_1=3) design target for the OPA637, the resulting key frequencies in the Bode analysis of **Figure 2**, and the component values necessary to set up this compensation. The selected feedback-resistor value is the result of a compromise between high input impedance ($R_G=R_F/(G_1-1)$) and keeping the compensation capacitors greater than the parasitic values on those nodes.

To implement this circuit for testing, you must also consider the components' parasitic capacitances and test-interface requirements. To include the effect of parasitics, the actual test-circuit design (**Figure 3**) reduces the value of C_F by 0.2 pF and that of C_S by the 15-pF parasitic at the input of the OPA637. The test circuit also includes 50Ω impedance-matching resistors at the input and output to match the assumed test-equipment source and load impedances of 50Ω.

Adding the input-matching resistor slightly changes G_1 from 3.0 to 2.95. This change has no effect on F_0 and very little effect on Q because the G_1Z_0 portion of **Equation 11** is small relative to GBP/G_2. The test circuit also shows a bias-current-cancellation resistor from the noninverting input to ground. This resistor is equal to the parallel value of R_F and R_G to improve the output dc offset that results from bias currents. With this resistor match in place, the output dc error that results from input bias currents is simply the input-offset current times the feedback-resistor value. A large capacitor shunts this noninverting-input resistor to roll off the noise terms that might arise from the resistor's Johnson noise and bias-current noise. These two components on the noninverting input are unnecessary for the FET-input OPA637 because its bias, offset, and noise-current terms are infinitesimal relative to the voltage offset and noise terms. The test circuit in **Figure 3** includes these components for general application.

The test circuit's frequency response (**Figure 4a**) is remarkably flat for a gain-of-5-stable op amp operating at a noise gain of 3 (a signal gain of –2). The frequency response does show slight peaking, which indicates that the Q of the actual circuit is slightly greater than 0.707 instead of the target value of 0.64. This difference in Q slightly extends the bandwidth from a target of 7.7 to 9.8 MHz and introduces some overshoot and ringing into the pulse response (**Figure 4b**). Apparently, either the target G_2 was too close to the minimum stable gain to exclude the effects of the higher order poles or the parasitic capacitances are different from the estimate. You can obtain a closer match between predicted and measured results at higher values of G_2.

You can compare this inverting compensation for the OPA637 to a maximum-bandwidth design using the unity-gain-stable OPA627 at a gain of 2. The inverting compensation with the OPA637 produces a slightly higher bandwidth of 9.8 MHz vs the 8 MHz of the OPA627. However, because of the slew-rate difference, the OPA627 is slew-limited for output steps greater than 2.4V, whereas the OPA637 can support nonslew-limited steps as high as 4.2V at the output. If this 4V were the input range of an ADC, the nonslew-limited pulse response that the OPA637 provides would settle to a final value more quickly than that of the OPA627.

For example, this compensation of the OPA637 provides a low-gain ADC buffer with excellent settling time for large output steps. When driving a $4V_{p\text{-}p}$ input-range, 10-bit ADC, the circuit has an absolute dc accuracy (with no trims) and peak-to-peak output noise that doesn't exceed ¼LSB. The worst-case output dc error is 0.75 mV, and the worst-case output peak-to-peak

TABLE 4—GAIN-BANDWIDTH-INDEPENDENT DESIGN

Design targets	Resulting design	Predicted Q and $F_{-3\,dB}$ vs gain			
		G_1	$-R_F/R_G$	Q	$F_{-3\,dB}$ (MHz)
F_0=5 MHz	Z_0=312 kHz	2	–1	0.776	5.59
Nominal G_1=4	G_2=13.74	3	–2	0.74	5.31
Nominal Q=0.707	P_1=4.3 MHz	4	–3	0.707	5
		5	–4	0.677	4.64
		6	–5	0.65	4.24
		7	–6	0.624	3.75
		8	–7	0.601	3.16

noise is 0.9 mV. The settling time to ½LSB is 33 nsec.

The improvement in performance between unity-gain-stable and decompensated versions of the same op amp is even more significant if you use parts that have a wider difference in their minimum stable gains.

Predict the output noise

Any compensation technique that shapes the noise gain of a nonunity-gain-stable op amp produces higher output noise as the frequency increases. This compensation technique increases the gain for the noninverting input-voltage noise of the op amp, as the noise-gain portion of **Figure 2**'s Bode plot shows. In most cases, the op amp's noninverting input-voltage noise dominates the total output noise for the circuit of **Figure 1**. Referring to the Bode plot, this input-voltage noise has a gain that starts at G_1, has a zero at Z_1 equal to G_1Z_0, and finally has second-order poles that are identical to those you set in the inverting-compensation design. The transfer function for either the noise or a signal applied to the noninverting input (V^+) of **Figure 1** is

$$\frac{V_O}{V^+}=\frac{GBP\,(s+G_1Z_0)}{s^2+s\frac{\omega_0}{Q}+\omega_0^2}. \qquad (18)$$

(Refer to **Equations** 8 and 9 for the terms that you can place in this equation.)

One approach to describing the output noise is to compute an equivalent noise-power bandwidth (NPB) that, when you multiply it by a constant output-noise-power value, gives the same total integrated noise power as the actual frequency response. If you arbitrarily use the output noise due to the noninverting input-voltage noise amplified by G_1, which is the low-frequency output noise due to the noninverting input-voltage noise, as the constant noise value, you can calculate an equivalent NPB as

$$NPB=\frac{1}{G_1^2}\int_0^{\infty}\frac{GBP^2(\omega^2+G_1Z_0^2)d\omega}{\omega^4+\omega^2\omega_0^2(\frac{1}{Q^2}-2)+\omega_0^4}. \qquad (19)$$

This noise is the square of **Equation 18**'s gain magnitude integrated from a frequency of 0 to infinity, then divided by the low-frequency noise gain squared (G_1^2). This integral simplifies considerably, and you can solve it in closed form when you target a Q of 0.707. The middle term in the denominator of **Equation 19** drops out, which allows you to use integral-table solutions for forms including $1/(x^4+c^4)$. Using the terms defined in **Figure 2** and assuming a Q of approximately 0.707 gives an equivalent NPB of

$$NPB=\frac{3\pi}{4\sqrt{2}}\left(\frac{GBP}{G_1}\right)^2\frac{1}{F_0}\left[1+\frac{G_1Z_0}{GBP/G_1}\right]. \qquad (20)$$

The last term in this **equation** is generally much less than 1, and you can ignore it. This **equation** states that the NPB is approximately equal to the single-pole bandwidth that results if you simply operate the amplifier at G_1 (a bandwidth equal to GBP/G_1) times the ratio of that bandwidth to the characteristic frequency ($F_0=\sqrt{Z_0GBP}$) of the actual second-order closed-loop response.

To use this calculated NPB, multiply the op amp's noninverting input-voltage noise by G_1 to compute the low-frequency spot noise at the output. Then, multiply that result by the square root of **Equation 20** to get the integrated noise ($E_{O(RMS)}$). Performing these computations for the design example of **Figure** 3, which has a Q≈0.707, gives

$$E_{O(RMS)}=\left(4.5\ nV/\sqrt{Hz}\right)3\left[\frac{3\pi}{4\sqrt{2}}\left(\frac{80}{3}\right)^2\frac{MHz}{8.6}\right]^{1/2} \qquad (21)$$
$$=\left(13.5\ nV/\sqrt{Hz}\right)\sqrt{137\ MHz}=158\ \mu V_{RMS}.$$

This analysis shows the significant increase in output-voltage noise due to the increased noise gain to G_2 by assuming a constant output noise and computing the required NPB to get the same integrated noise power as the actual output noise over frequency. Evaluating this integral for the NPB is based on the assumption that the op amp's frequency response of **Equation 18** self-limits the output noise.

Another way to look at this noise is to compute the equivalent input-voltage noise that integrates to the same power over a simple lowpass Butterworth bandwidth. This approach allows an easy comparison between this technique and other approaches for getting a desired frequency response. The NPB of a simple, second-order Butterworth response equals $1.11F_0=1.11F_{-3\ dB}$ when Q=0.707 (see **box**). You can set up an equality to define the equivalent input-spot noise (E_M) that will integrate over an NPB set by $1.11F_0$ to the same total output-noise power as the actual response as follows:

$$E_NG_1\sqrt{\frac{3\pi}{4\sqrt{2}}\frac{(GBP/G_1)^2}{F_0}}=E_MG_1\sqrt{1.11F_0}, \qquad (22)$$

where E_N is the input-voltage noise of the op amp. The solution for E_M is

$$E_M=E_N\sqrt{1.5\bullet\frac{GBP/G_1}{Z_1}}. \qquad (23)$$

This **equation** states that the increase in the equivalent input-referred spot-noise voltage is proportional to the square root of the ratio of GBP/G_1 to Z_1. Evaluating **this equation** for the design example in **Figure 3** yields an equivalent input-noise voltage of 17.1 nV/√Hz. Multiplying this result by the low-frequency noise gain, G_1, and then by the square root of $1.11F_0$ (**Table 2**) gives the integrated noise. This calculation gives the same 158 µV of integrated noise as **Equation 21** gives. **Equation 23** is useful because it clearly shows

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